REMOTE SENSING LABORATORY

(12)

13.9 GHZ RADIOMETER DESIGN

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ABSTRACT

This report concerns the design of a 13.9 GHz total power radiometer. A large amount of literature is available on the subject, but this report addresses a detailed design concept without going into the mathematical analysis. All through the discussion, contrast is brought forth between a radar and radiometer, since the former is a more familiar system to most of the beginners in microwave system design. No stress is laid on the theoretical aspects of the radiometer.

The system described has an overall noise figure of 7.67 dB and uses an integration time of about 1 sec. It is capable of a resolution of 2 K and the slope of the transfer function is set as 1 mV/1 K for convenience.

1.0 INTRODUCTION

1.1 Need for A Radiometer

Radiation emitted as well as reflected by any object is known to be a function of various physical properties of the object itself; hence, measured radiation data can be used to understand the properties of the scene under observation. The well-established science of measuring such radiation is called "radiometry" and the device used to carry out the measurements is known as a "radiometer". This report describes the systematic design and calibration procedure of a 13.9 GHz practical radiometer used to probe sea ice.

For convenience, the report is organized into six sections and various sub-sections. In a broad sense, Section 1 describes the overall system. Section 2 illustrates the practical design with some analysis based on the references. Section 3 is used to bring out some of the methods of implementing the most-often required system measurements. Calibration of the system is treated in adequate detail in Section 4, followed by the description of operation in Section 5. Since many of the components used in this system fell short of the requirements, Section 6 is used for suggestions to improve the performance of the system. Specifications and the measured data are presented in Appendix A. Appendix B gives the FORTRAN program listing used for computing the calibration curves.

1.2 Why Is It Different From a Radar Receiver

Unlike a radar, a radiometer is a passive sensor. Another major difference between a radar and radiometer lies in the fact that the unwanted noise present at the input of the receiver antenna of a radar is the signal of interest in the case of a radiometer. On paper, a radiometer looks like any conventional superheterodyne receiver, but in practice the design approach is

quite different, as will be evident in a later section on design. Basically, radiometers are divided into one of two types: (a) total power mode or (b) switched mode, the latter being powerful. The type discussed here operates in the total power mode, but can be modified to work in the switched mode.

1.3 Description of a 13.9 GHz Total Power Radiometer

Figure 1 shows the block diagram of the complete system. The antenna used is a standard gain horn connected to one of the four ports of a waveguide microwave switch. The receiver and noise diode are connected to the other two ports and the last port is left open. During measurements, the connections at

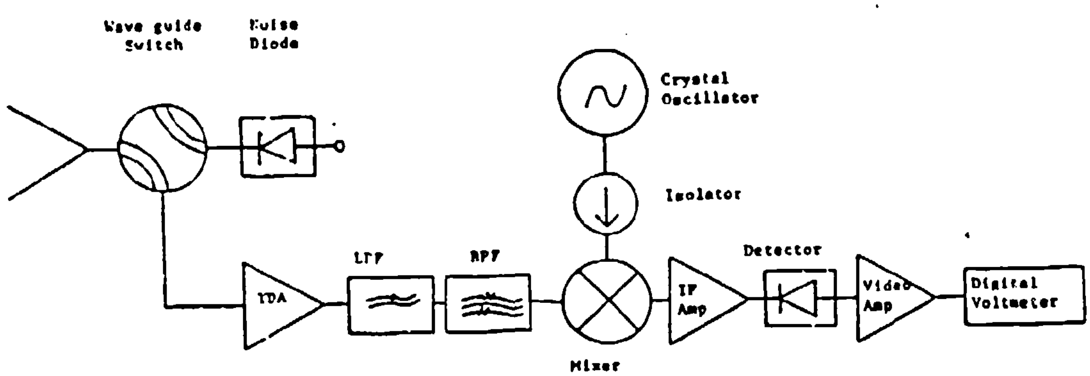


FIGURE 1: Total Power Radiometer Block Diagram

the switch will be just as shown in Figure 1, i.e., the antenna is coupled to the receiver and the diode is coupled to the open port. In this condition, the switch is said to be in position 1. During calibration (position 2), however, the antenna is coupled to the open port and the diode to the receiver to deliver a known amount of noise. The switching between positions 1 and 2 is done by applying +28 VDC to the switch. Since the isolation between the two guides of the switch is typically 60 dB, the signals (actually noise) in the two guides do not interact very much.

The signal is amplified in the tunnel diode amplifier and band-limited in a narrow-band filter with good (> 50 dB) stop-band rejection. In practice, the filtering is done in two parts, the first one being a lowpass filter and the other a bandpass filter. The bandpass filter is a four-cavity tunable filter and it may also respond to the harmonics of the signal (i.e., 27.8 GHz, 41.7 GHz, etc.). The lowpass filter is used to suppress these harmonics. The double-balanced mixer used has a waveguide RF port and LO and IF ports in SMA connectors. The local oscillator used for superheterodyne mixing is a 13.93 GHz crystal-stabilized source. The mechanism behind the generation of the 13.93 GHz signal involves a fundamental oscillator at 1.393 GHz which will be multiplied by x10 to get the required frequency. Phase locking is achieved by a reference crystal oscillator operating at 100 MHz which, in turn, will be fed to a harmonic generator whose output will be compared with the 1.393 GHz fundamental in a phase detector. The error signal output of the phase detector is used to tune the fundamental oscillator. The isolator provides a better match between the oscillator and the mixer. The intermediate frequency (IF) amplifier is a high-gain discrete version operating at 30 MHz frequency with a 3 dB bandwidth of only 1.2 MHz. As will become evident in later discussion this small bandwidth is a serious limitation to the feasible system resolution. Square-law detection is carried out by a crystal detector with reasonably good sensitivity. At this point, it is interesting to note that the expected change in signal level at the detector output over the entire dynamic range of temperature is just a fraction of a millivolt. This calls for a large video amplification to scale up the signal to a reasonable value, say, a few hundred millivolts. The resulting signal is integrated for a suitable period of time (about a second) and read out on a digital voltmeter.

1.4 How to Convert It Into a Switching Type

To convert the existing system into a switching-type radiometer the RF switch must be switched on and off continuously between the antenna and the reference load (the diode noise source in this case). Usually a square wave is used for switching, thus spending 50 percent of the time in the desired signal measurement and the other 50 percent in calibration. The switching frequency is limited by the RF switch response. Also, the RF detected signal must be bandpass-filtered at both the switching frequency and some of its harmonics. As a rule of thumb, at least two harmonics must be included, and the signal should be synchronously detected before integration.

All of the required microwave components were available in the lab, but not their specifications. Therefore, every component was evaluated and tuned wherever necessary to meet the required specifications as far as possible. The results of the measurements are shown in Appendix A.

2.0 SYSTEM DESIGN

2.1 Necessary Equations

In essence, a radiometer must exhibit a linear transfer characteristic between the input temperature and the output voltage. Since the input power is proportional to the temperature, it also has a linear relationship between the input power and the output voltage. Mathematically, the input power to the system, as referred to the antenna terminals P_{sys}, is given by

$$P_{sys} = k\ T_{sys}\ B \tag{1}$$

where:

k = Boltzman's constant

B = effective bandwidth of the IF amplifier (Hz)

T_{sys} = system temperature referred to the antenna terminals (K)

In turn, the system temperature can be expressed as:

$$T_{sys} = \eta T_a + (1-\eta)T_p + (L-1)T_p + LT_{rec} \quad (2)$$

where:

η = antenna radiation efficiency

L = transmission loss between the antenna and the receiver

T_a = antenna temperature of the scene under observation

T_p = physical temperature of the system

T_{rec} = receiver noise equivalent temperature.

A detailed derivation of these equations can be found in many books (e.g., Ulaby, Moore and Fung, 1981).

2.2 Set the Specifications

Once the input and output relations are known, the first thing towards the system design is to define the following system specifications.

1. The dynamic range of the system, which usually means the difference in power between the maximum signal (the fact that the receiver may saturate at higher levels leading to an undesirable nonlinear transfer function sets the limits) and the minimum detectable signal (usually set by the thermal noise floor) expressed in dB. However, for a radiometer's system temperature (linearly related to the power for a fixed bandwidth as long as the system is maintained at a constant temperature), it may be more realistic to express the system dynamic range.

2. The required output voltage range corresponding to the expected input temperature range, which is the slope of the system transfer function.
3. Overall system noise figure, which is the one that sets the thermal noise floor mentioned in 1 above.
4. Bandwidth of the intermediate frequency amplifier.
5. Fractional gain variation of the overall system.
6. The dynamic range limits of the square-law detector.

A detailed consideration of these parameters is applied to the system under discussion. Since the radiometer is a passive sensor and the noise signals available are very small, the upper limit of the dynamic range is not a problem. In this particular case, since the system is expected to operate at very close ranges (as low as a few meters), the lower limit also does not pose any real problem. In medical applications it is more of a problem because the antenna will be always in the near field. In the case of radiometers used in astronomy, where high sensitivity is of paramount importance, the lower limit plays a critical role. For this radiometer a 20K - 280K input range is set and the output voltages corresponding to this are required to fall between 0 mV and 260 mV, which means a slope of 1 mV/1 K. Even though the range of temperatures expected in the case of sea ice is much smaller, the above specification is set in view of the calibration (which will be discussed elsewhere).

The measured overall system noise figure with the available components F_n = 7.67 dB = 5.85. From this the receiver noise equivalent temperature T_{rec} is calculated as:

$$T_{rec} = (Fn - 1)\, T_0 = (5.85 - 1)\, 273 = 1324.05K$$

The gain variation $\Delta G/G$ of the complete front end is assumed to be equal to 10^{-3}. In practice,it is somewhat difficult to measure this variation so a setup with a stability of at least an order better is needed. For this discussion the value assumed above is believed to be reasonable.

3 dB bandwidth of the IF amplifier available = 1.2 MHz.

Dynamic range of the square-law detector = 10 dBm-(-50 dBm) = 60 dB.

Other pertinent data for individual components is shown in Appendix A.

2.3 Practical Design

2.3.1 Front End

Using Equation (2) in the middle of the dynamic range, the system temperature T_{sys} can be calculated as:

$$T_a = 150 \text{ K}$$

$$T_{sys} = 0.7 \times 150 + (1-0.7)273 + (1.02-1)273 + 1.02 \times 1324.05 = 1542.9 \text{ K}$$

Similarly,

minimum T_{sys} = 1451.9 K

and

maximum T_{sys} = 1633.9 K.

2.3.2 A Look at the Resolution

The minimum detectable change in antenna temperature known as radiometric resolution ΔT, is given by

$$\Delta T = T_{sys} \left[\frac{1}{Bt} + \left(\frac{\Delta G}{G}\right)^2\right]^{1/2} \quad (3)$$

The derivation of this equation can be found in Ulaby, Moore and Fung [1981] and Evans and McLeish [1977]. The worst case of resolution occurs for maximum T_{sys} and, hence, it is preferable to use this value in the above formula. For different values of integration time the calculated radiometric resolution is shown in Table I as well as in Figures 2 and 3.

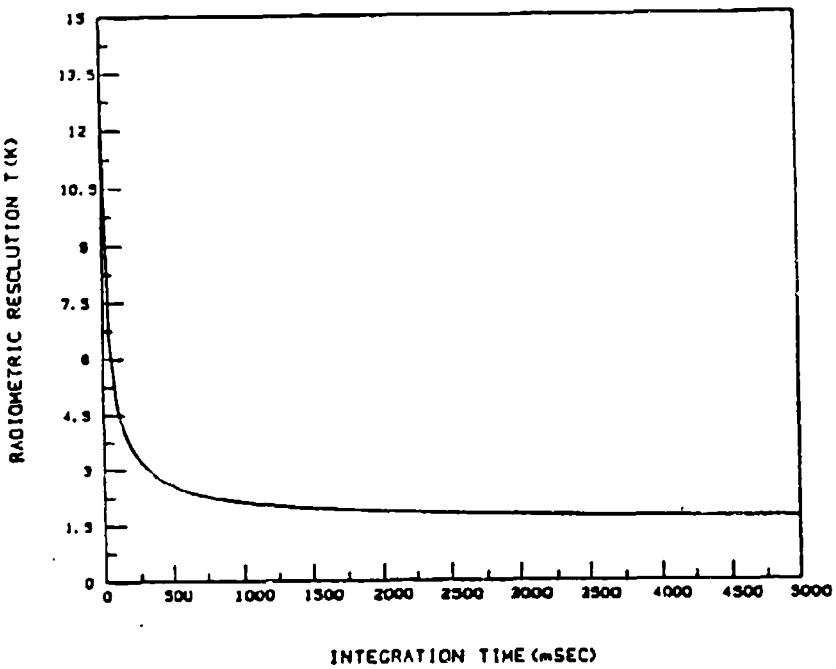


FIGURE 2: Radiometric Resolution vs Integration Time for $\Delta G/G$ = .001

Equation (3) shows that, for a given system temperature and fixed bandwidth, the radiometric resolution improves as the square root of the integration time, provided the fractional gain variation is low. Table I shows that the resolution saturates for $\Delta G/G > 1/Bt$. Hence, it is very important that the fractional gain variation be kept as low as possible to obtain fine resolution for a reasonable value of the time-bandwidth product. Fine-resolution systems usually achieve that kind of resolution by using large

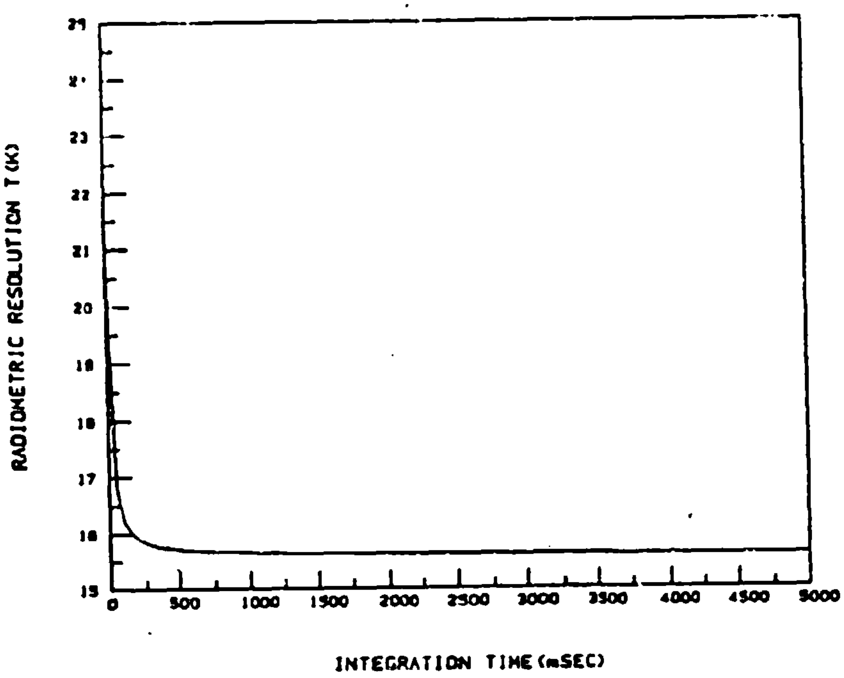


FIGURE 3: Radiometric Resolution vs Integration Time for $\Delta G/G = .01$

TABLE I

Integration time t (msec)	Radiometric resolution ΔT (K)	
	$\Delta G/G = 10^{-3}$	$\Delta G/G = 10^{-2}$
1	44.96	47.57
10	14.29	21.08
100	4.76	16.21
1000	2.12	15.64
2000	1.85	15.60
5000	1.69	15.59
10000	1.62	15.57

bandwidths rather than large integration time. Also, since the effective noise bandwidth is greater than the 3 dB bandwidth of the IF amplifier, the actual resolution in each of the entries in Table I will be better (less) than the values shown. For the current system, since the integration time is about 1 sec, the radiometric resolution can be expected to be approximately 2 K.

As can be seen from equation (3), the radiometric resolution ΔT is proportional to the system temperature when all other quantities in the equation are fixed. The T_{sys}, in turn, depends upon the antenna temperature, which is related to the temperature of the scene under observation and the receiver noise-equivalent temperature T_{rec}. For fixed T_a, the resolution ΔT varies in proportion to the receiver noise-equivalent temperature. This is where the importance of the low-noise-figure requirement for receivers (and hence the use of an expensive low-noise amplifier) comes into the picture.

In this case for a $\Delta G/G$ of 10^{-3} the effect of receiver noise figure on resolution with integration time as the parameter is calculated. The results are shown in Table II and also plotted in Figures 4-7.

TABLE II

Receiver NF	Radiometric Resolution ΔT (K)			
(dB)	t = 1 msec	t = 10 msec	t = 100 msec	t = 1 sec
2	12.90	4.10	1.36	0.60
3	16.19	5.15	1.71	0.76
4	20.35	6.47	2.15	0.95
5	25.58	8.13	2.71	1.20
6	32.16	10.23	3.40	1.51
7	40.45	12.86	4.28	1.90
8	50.89	16.18	5.38	2.39
9	64.03	20.36	6.77	3.00
10	80.57	25.62	8.52	3.78

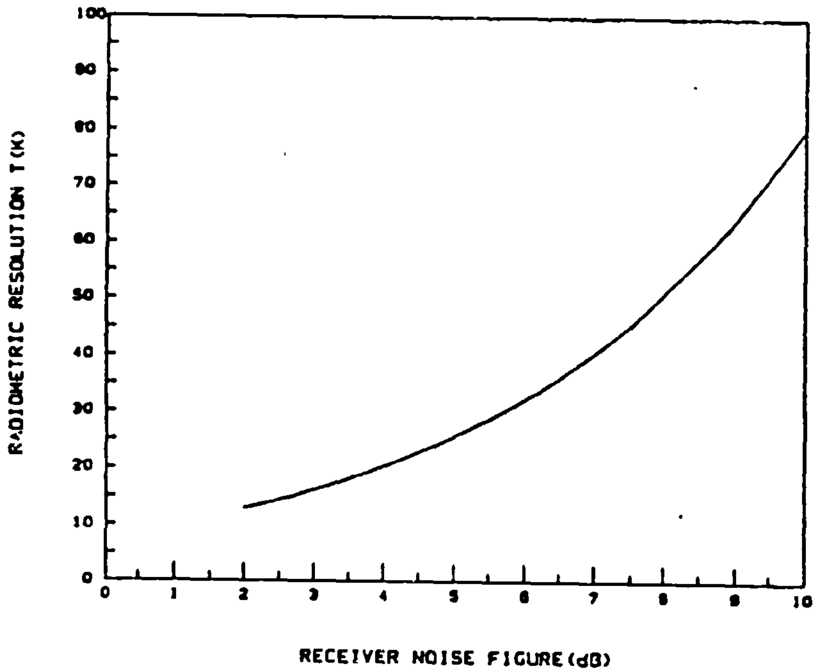


FIGURE 4: Radiometric Resolution vs Receiver Noise Figure for t = 1 mSEC

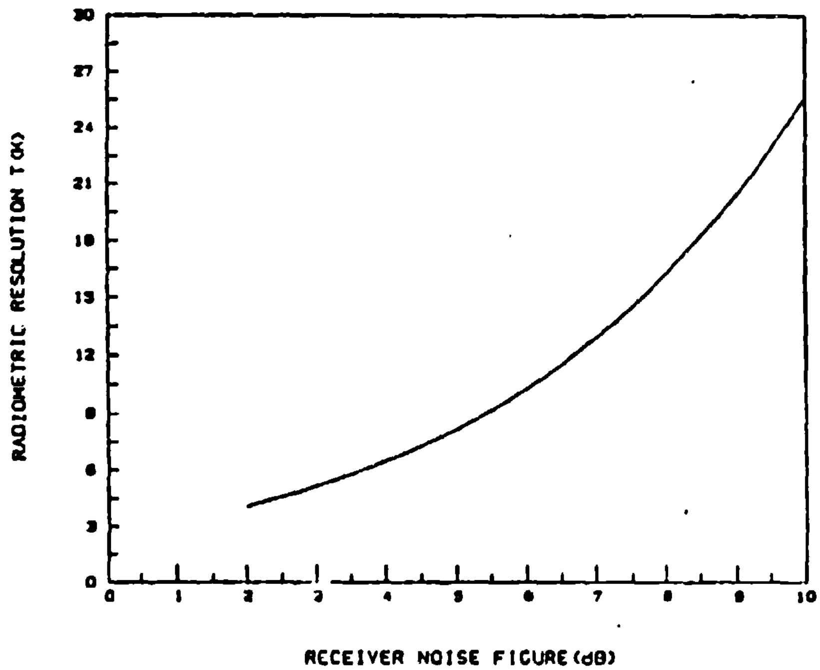


FIGURE 5: Radiometric Resolution vs Receiver Noise Figure for t = 10 mSEC

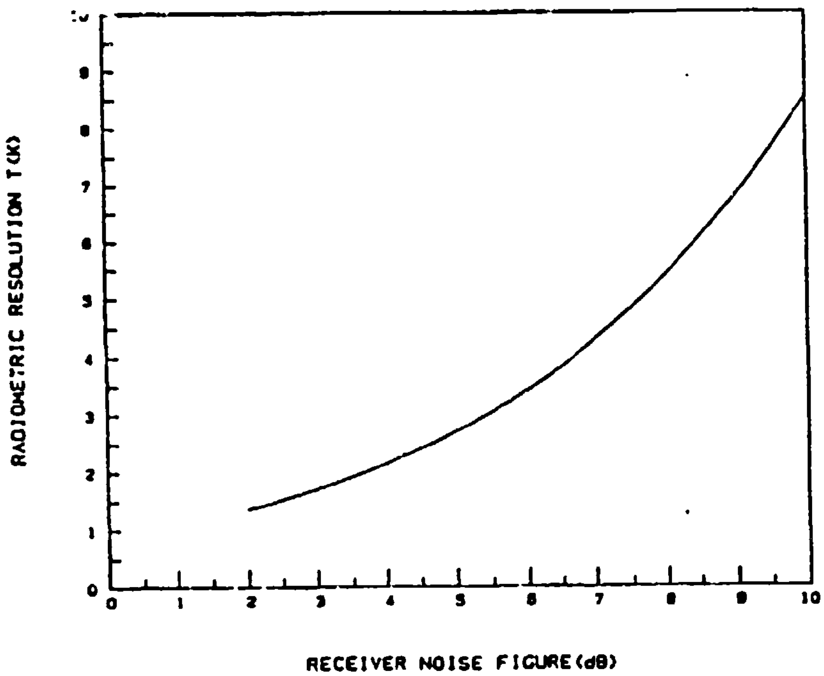


FIGURE 6: Radiometric Resolution vs Receiver Noise Figure for t = 100 mSEC

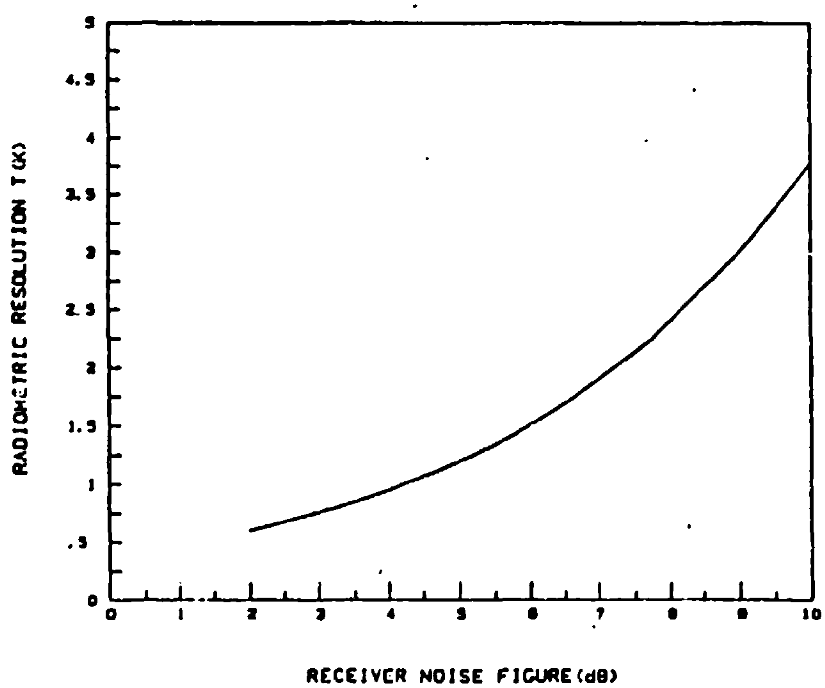


FIGURE 7: Radiometric Resolution vs Receiver Noise Figure for t = 1 SEC

Table III and Figure 8 show the dependence of radiometric resolution ΔT on the noise bandwidth, provided all other quantities are held constant. The plot is calculated from equation (3) using maximum T_{sys}, t of 1 sec, and $\Delta G/G = 10^{-2}$ and 10^{-3}.

TABLE III

Bandwidth B (MHz)	ΔT (K) with $\Delta G/G = 10^{-3}$	ΔT (K) with $\Delta G/G = 10^{-2}$
1	2.20	15.65
2	1.91	15.61
3	1.80	15.60
4	1.74	
5	1.71	
6	1.68	
7	1.66	
8	1.64	
9	1.63	
10	1.61	

This indicates that no matter how much bandwidth and "t" you use, a gain variation of 10^{-2} is totally unacceptable. In fact, for a 1 K resolution and a feasible integration time of 1 sec, $\Delta G/G$ needs to be much better than 10^{-3} (maybe 10^{-4}). The other alternative is to use a switching mode radiometer.

2.3.3 IF Section and Detection

The next step in the design process is to calculate the required gain for the complete front-end system (i.e., up to the square-law detector). The criterion, as mentioned earlier, is to see that the signal (noise) power corresponding to the mid-point of the temperature dynamic range coincides with the

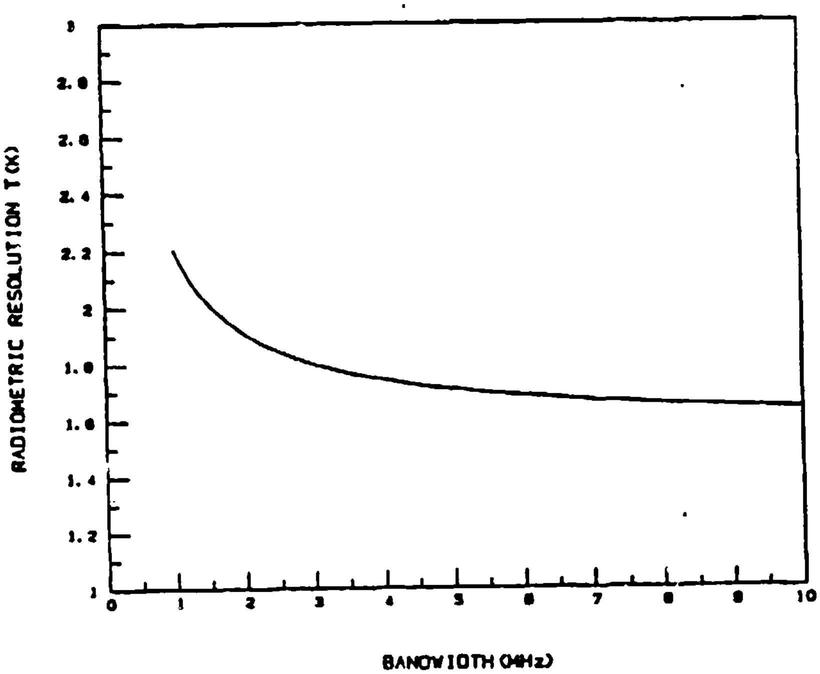


FIGURE 8: Radiometric Resolution vs Bandwidth for $\Delta G/G = .001$

mid-point of the square-law detector dynamic range. It should be noted that, if no other component prior to the detector is driven in to saturation, the dynamic range is going to be limited by the detector. However, in radiometers this situation seems almost impossible as long as the above-mentioned criterion is satisfied. In the present case, the ratio of the maximum to the minimum system temperature is much less than 2, indicating that we need less than 3 dB of the detector's available dynamic range. Nevertheless, it is good practice to position the signals at the center of the response of the square-law detector. In this case an antenna temperature of 130 K, corresponding to a system temperature of 843.88 K, must produce a -20 dBm (10^{-5} W) signal at the input of the square-law detector. The receiver gain G_{rf} required to satisfy this condition would be calculated as follows:

$$G_{rf} = \frac{P_{if}}{K\ T_{sys}\ B\ F_n} \qquad (4)$$

where:

G_{rf} = receiver power gain (without antenna) in dB

P_{if} = power output of the IF amplifier corresponding to the T_{sys}.

$$G_{rf} = \frac{10^{-5}}{1.3805\times10^{-23} \times 1542.9 \times 1.2\times10^{+6} \times 5.85} = 6.69\times10^{+7} = 78.25 \text{ dB}$$

Note that if the actual effective noise bandwidth, rather than the 3 dB IF bandwidth, is used in the expression, G_{rf} may decrease about a dB or so. For this system, the tunnel diode amplifier has a gain of 16 dB and the single-side band conversion loss of the mixer is 21.2 dB. The pre-selection filter insertion loss is 0.3 dB and the waveguide loss between the amplifier and the switch is 0.1 dB. Therefore, the required IF amplifier gain G_{if} is given by:

$$G_{if} = G_{rf} - 16 + 21.2 + 0.3 + 0.1 = 83.85 \text{ dB.}$$

A crystal detector of sensitivity, r, 500 mV/mW is used for square-law detection. The sensitivity of such a detector is expressed in millivolts/millwatts because the input power is linearly related to the output voltage, not to the output power. From this information the minimum and maximum values of the voltages available at the input of the video amplifier are calculated as:

$$V_{in} = r \times K \times T_{sys} \times B \times F_n \times G_{rf} \qquad (5)$$

$$\text{Minimum } V_{in} = 500\times1.3805\times10^{-23}\times1451.9\times1.2\times10^{+6}\times5.85\times6.69\times10^{+7} = 4.707 \text{ mV}$$

$$\text{Maximum } V_{in} = 500 \times 1.3808 \times 10^{-23} \times 1633.9 \times 1.2 \times 10^{+6} \times 5.85 \times 6.69 \times 10^{+7} = 5.297 \text{ mV}$$

2.3.4 Back End

Notice that the change in the detector output voltage is just 0.32 mV for a change of 260 K at the input of the antenna. This calls for a large video amplification to scale up the small voltage change to a sufficiently large change. Thus the video amplifier gain G_v is computed as:

$$G_v = \frac{\text{desired output slope}}{\text{available input slope}}$$

$$= \frac{1 \text{ mV/K}}{(5.297 \text{ mV} - 4.707 \text{ mV})/(280 \text{ K} - 20 \text{ K})} = 440.678$$

$$G_v = 20 \log (440.68) = 52.88 \text{ dB.}$$

Thus at the output of the video amplifier the voltage levels are:

$20 \text{ K} \rightarrow 2.074 \text{ V}$

$280 \text{ K} \rightarrow 2.334 \text{ V}$

Finally, by providing an offset voltage of 2.074 V, we get

$20 \text{ K} \rightarrow 0.0 \text{ mV}$

$280 \text{ K} \rightarrow 260 \text{ mV}$

Another requirement is that the video amplifier must provide proper terminating resistance for the detector. In this case it should be approximately 1 K. Also, because of the dc component present (the actual signal is not dc, but a very low frequency which will be further integrated) at the output of the detector, the final offset might be somewhat different from the calculated 1.089 V value. Thus it is advisable to provide the offset after integration rather than before.

As mentioned above, all of the front end components were available in the lab. The measured data for each of these components are listed in Appendix A. The complete back end of the radiometer is built as a single module. The circuit diagram is shown in Figure 9. It consists of a high gain (approximately 60 dB) video amplifier IC-1, lowpass filter IC-2, and an offset circuit IC-3. An instrumentation amplifier, rather than discrete operational amplifiers, is chosen for video amplification. It not only reduces the chip count but also eliminates any possibility of small oscillations associated with high gain amplification (with high gain even a small positive feedback is enough to initiate oscillations). These oscillations may sometimes be very close to the desired signal frequency, making it impossible to eliminate even by lowpass filtering. Since an instrumentation amplifier is internally compensated, all of these problems will automatically be taken care of. An additional advantage of the instrumentation amplifier is the fact that the IC has an on-chip common point (operational amplifiers do not have this) making it less susceotible to spurious pickups. As shown in Figure 9, potentiometer R1 is used to adjust the offset of the instrumentation amplifier and R2 to trim its gain. The input is terminated in 1 K ohms to provide matching to the output of the square law detector. The integrator or lowpass filter is built with a BIFET

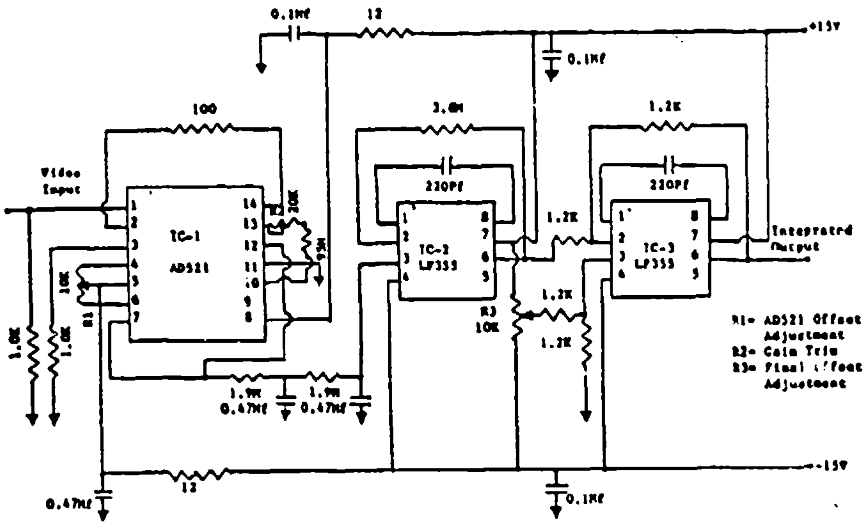


FIGURE 9: Video Amplifier Circuit Diagram

operational amplifier (LF 355) and has a time constant of RC - $1.9^{+6} \times 0.47^{-6}$ = 0.893 seconds. It is a two-pole Butterworth filter. Finally, the offset circuit is realized with a BIFET as a differential unity gain amplifier. Potentiometer R3 is used to adjust the final offset voltage. Thus, the output of this circuit is the required integrated signal and read on a digital voltmeter.

All the modules are mounted on a 2-mm-thick aluminum plate. The layout of the various modules and interconnections is shown schematically in Figure 10. Also shown are the dc power supplies required for different modules.

3.0 SYSTEM MEASUREMENTS

Two of the most important specifications of any microwave reciever system are power gain and noise figure. Each component of the system will affect, to some extent, one or both of these quantities. Hence, proper measurement of

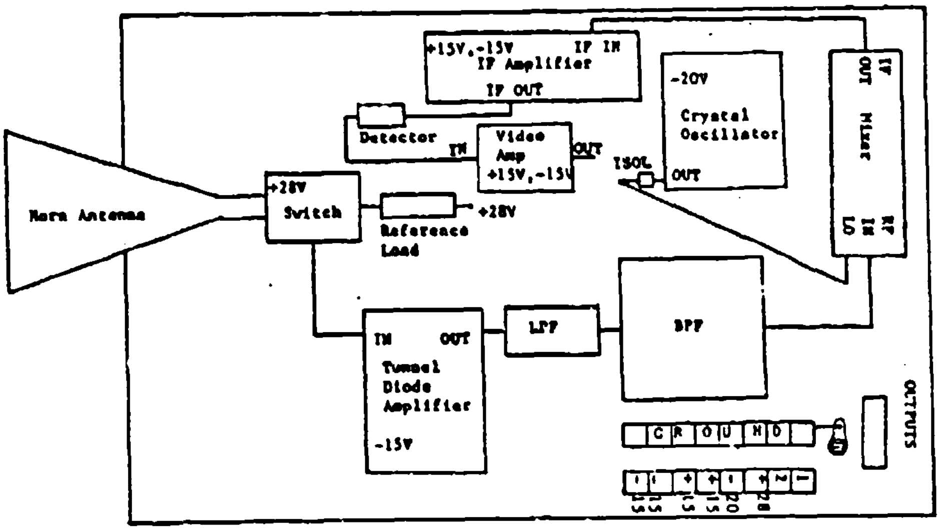


FIGURE 10: Layout of the Various Modules

these quantities for each of the system components (to make required adjustments to the system) and, finally, the system (to evaluate the system performance) is warranted. Usually, at microwave frequencies three types of measurement techniques are available: (1) manual; (2) semi-automatic; (3) automatic.

For single-frequency measurements, manual techniques are quite adequate; semi-automatic techniques are better suited for swept-frequency measurements. If high accuracy is required, automatic techniques are used. The last two are, naturally, much faster and often much easier to use. In this case, it is a single-frequency measurement and manual techniques are employed.

3.1 Power Gain or Loss

In one of the simplest setups (Figure 11) a signal source and a power meter are used. As usual, some interconnecting cables and adaptors are required. At the frequency of interest (13.9 GHz) the signal input to the device under test (DUT) in dBm is measured by bypassing the DUT as shown.

Next, the output of the DUT is noted down in dBm. The difference in the two readings gives the gain or loss in dB. There are many drawbacks to this setup, but the one of most concern is the mismatch between any two components of the setup -- source and DUT, DUT and power meter. The reflections caused by these mismatches result in inaccurate readings. Between the source and the DUT an isolator could be used (VHF through microwaves) but between the DUT and the power meter it may not be possible because, if the DUT is the complete receiver and the output is measured at the IF stage, no isolators may be available. Quite often, however, some amount of attenuation is mandatory between the DUT and the power meter in order to protect the power meter from being damaged by the large outputs of the DUT (Figure 12). This automatically takes care of the mismatch problem to some extent. Instead of an isolator, an attenuator could be used as long as the signal level is within the specifications of the DUT. In addition, a directional coupler of known coupling is used to measure the input signal to the DUT without disconnecting it from the setup. The amount of the input attenuation required is decided by the level of matching desired. For example, if there is a VSWR of 2.0:1 between the source and the DUT and a VSWR of 1.2:1 is desirable, the needed attenuation is half the dB difference betwen the two return losses -- one corresponding to each VSWR. So, magnitude of reflection coefficient

$$\text{without the attenuator} = (2-1)/(2+1) = 0.33$$

$$\text{return loss} = -20 \log (0.33) = 9.54 \text{ dB}$$

magnitude of reflection coefficient

$$\text{with the attenuator} = (1.2-1)/(1.2+1) = 0.09$$

$$\text{return loss} = -20 \log (0.09) = 20.9 \text{ dB.}$$

Therefore the minimum required attenuation = (20.9-9.54)/2 = 5.68 dB.

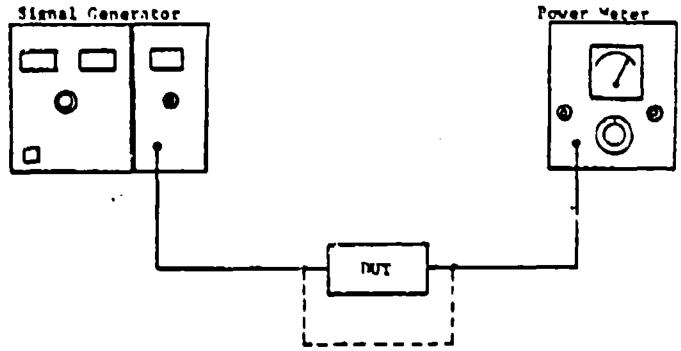


FIGURE 11: Basic Setup for Power Gain Measurement

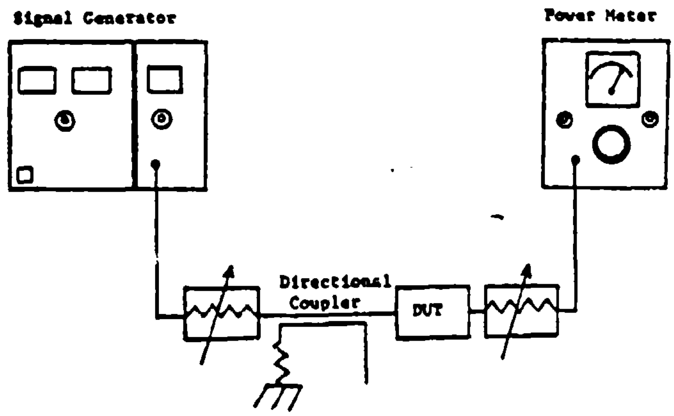


FIGURE 12: Improved Setup for Power Gain Measurement

A rigorous analysis (including the multiple reflections) yields a slightly larger value. Sometimes a variable attenuator is required to adjust the input power level to the DUT in which case the mismatch problem is automatically taken care of.

3.2 Noise Figure

The method described here is known as the Y-factor method and this is, by far, the most often used. One way of implementation is shown in Figure 13. Noise generated by a diode noise source of known excess-noise ratio (ENR) is fed to the DUT. An isolator of known insertion loss is used to reduce the mismatch. Once again, an attenuator with low VSWR could be used instead. The IF amplifier (30 MHz or 60 MHz) is not required if the DUT is the complete receiver and DUT output is the IF amplifier output of the receiver. The variable attenuator is used to adjust the power level and also to reduce the mismatch. The power meter may have to be very sensitive (the one used here has a range of -60 dBm to -20 dBm).

When the diode is off (cold condition - 0 VDC) the output power is noted in dBm as N1 and the diode is switched on (hot condition - +28 VDC). The output is noted again in dBm as N2. Now the noise figure in dB is calculated as [Laverghetta, 1981]:

$$NF = ENR\ (dB) - \text{Insertion loss of the isolator} - 10 \log \left[\text{Antilog}\ (N2/N1/10) - 1\right] \quad dB$$

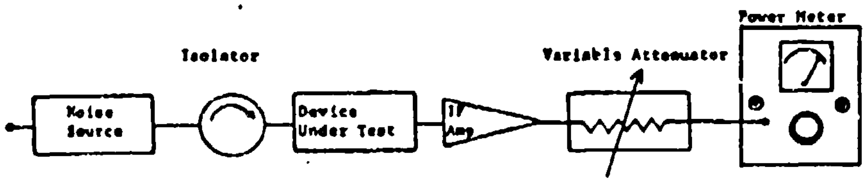


FIGURE 13: Setup Noise Figure Measurement

If an attenuator is used in place of the isolator its attenuation must be used in place of the isolator insertion loss in the above equation. The contribution of the external IF amplifier could be removed from the DUT noise figure if the IF amplifier noise figure and the DUT gain are predetermined.

4.0 CALIBRATION

4.1 Different Methods

Like any other linear system, radiometer calibration involves determining the relation between the output and input, in other words the system transfer function. In the case of a radiometer, the output parameter is the voltage measured after integration and the input parameter is the system temperature. Since the system is linear, at least two coordinates on the transfer characteristic must be found in order to complete the straight line. The source of the input must be stable and its antenna temperature be known quite accurately. This is very important because the whole system's performance depends on this. It may also be a good idea to choose the two coordinates so that each one is at a different edge of the transfer characteristic. There are many alternatives that can satisfy these requirements. Some of them are:

1. Clear (scatter free) sky
2. Eccosorb
3. Still water
4. Noise diode.

4.2 How to Do in Practice

The reference sources 1, 2 and 4 are quite dependable calibration sources. For a given frequency, the antenna temperature of the sky is a function of the angle of observation, atmospheric quantities (pressure, temperature, water vapor density profiles), and the presence of scattering media like clouds. For a clear sky condition all of the parameters above can either be measured or estimated with good accuracy. Therefore, this is one of the most commonly used reference loads in radiometry. Depending on the various parameters discussed above, the sky temperature may be anywhere between 2 K and 30 K. A FORTRAN program to calculate this temperature with different inputs is available on the Honeywell. For reference, the listing is given in Appendix B. The graph in Figure 14 is obtained for standard atmospheric profiles using this computer program. To obtain the first coordinate, point the radiometer antenna towards the sky at some convenient angle (small angles from zenith are preferable because the sky temperature doesn't vary too much). Make sure there are no clouds in the sky and note the output voltage. The other coordinate may be obtained by blocking the antenna aperture with eccosorb, a microwave absorbing material which is close to perfect (i.e., 100%) at these frequencies; it is well known that a perfect absorber exhibits an emissivity of unity. This means the antenna temperature will be the same as its physical temperature as long as the absorber is kept close to the antenna. The physical temperature of the absorber can be measured accurately by embedding a thermocouple (a metallic heterojunction which develops a voltage across the junction proportional to the junction temperature) in the absorber. The output voltage measured in this condition gives the second coordinate. The straight line joining these two coordinates (Figure 15) can be used to relate any output voltage to the corresponding system temperature.

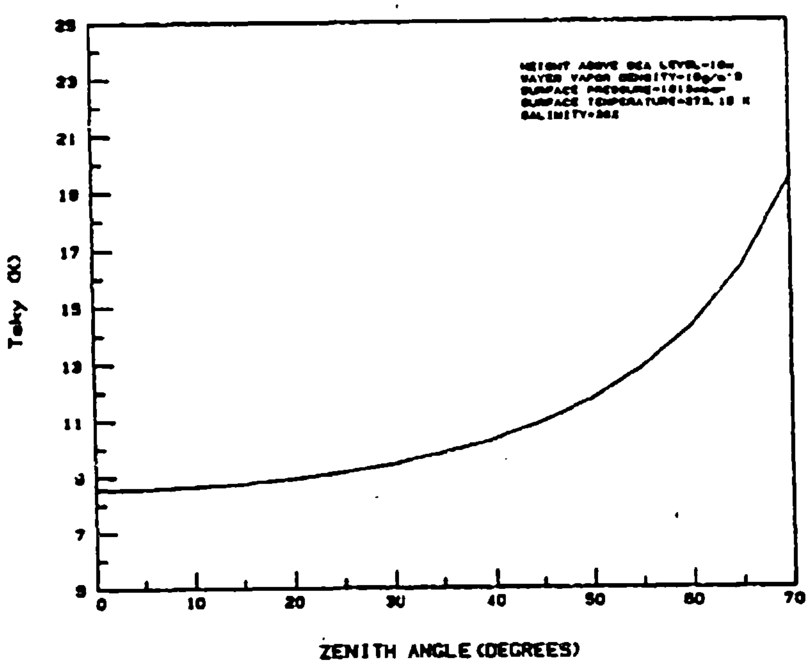


FIGURE 14: Sky Temperature vs Zenith Angle at 13.9 GHz

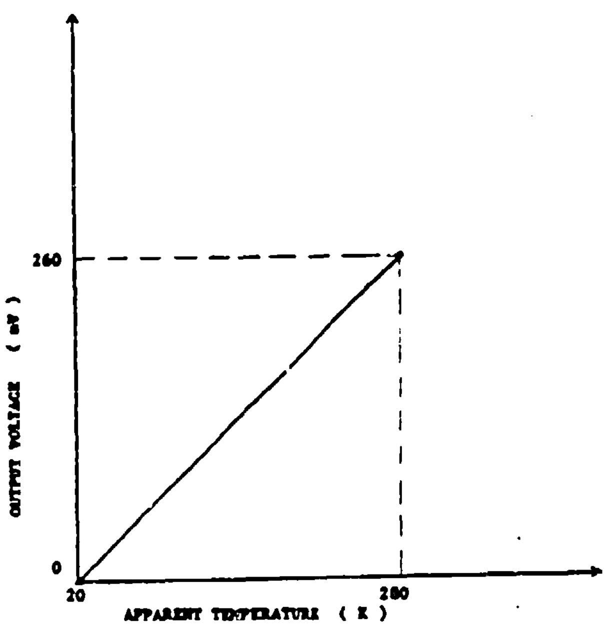


FIGURE 15: System Transfer Function

Still water is another useful reference source if the properties of the water (dielectric constant, physical temperature, foam coverage, etc.) are known accurately. Also required is the sky temperature (discussed above). The computer program mentioned before can also be used to calculate the antenna temperature of still water [Ulaby, Moore and Fung, 1984]. Note this temperature is also polarization dependent. Figure 16 shows the plots obtained using the computer program for vertical as well as horizontal polarizations. Once again, small angles measured from nadir seem to be better suited for calibration.

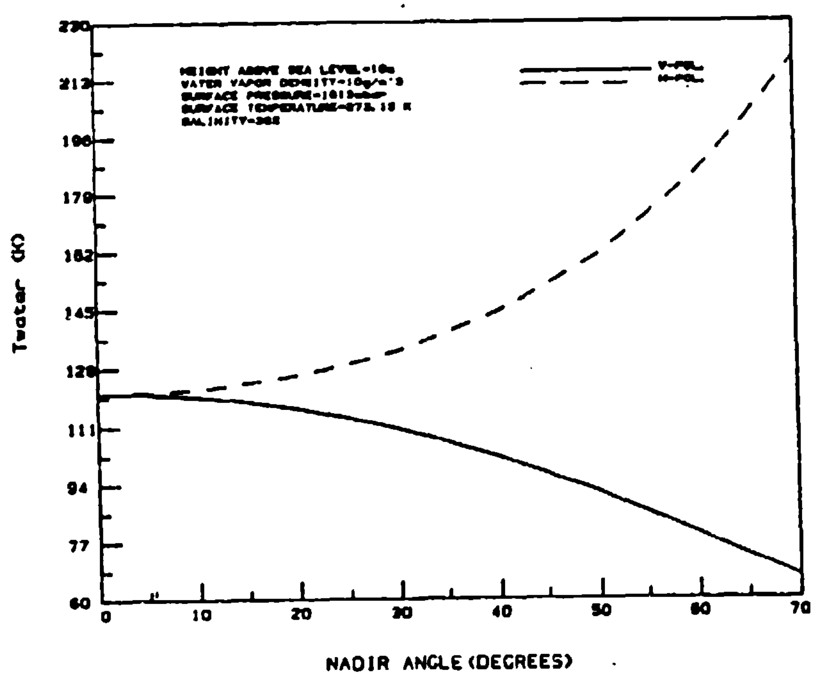


FIGURE 16: Water Temperature vs Nadir Angle at 13.9 GHz

Reference 4 has some advantages over 1 and 3 in the sense that it is not atmospheric dependent. In this case a diode capable of producing very wide band noise is turned "on" and "off" with a +28 VDC supply. If the noise power generated by this diode in one condition (on or off) is known, the noise power in the other condition can be calculated provided the excess noise ratio of the diode is known. In turn, it can be related to temperature and, hence, can be used for calibration. This type of reference is ideal for switching types of radiometers. Also, it may be noted that by calibrating the instrument at regular intervals, the effect of the system gain variations on the desired output can be minimized.

5.0 OPERATION

Operation of this system is fairly simple. Like many other systems, this radiometer needs some warm-up. Therefore, switch on the system and keep on standby for at least 10 minutes. This assures a stable local oscillator output within the specified limits and also minimizes drifts in other circuits. If the RF switch is not already in position 2, apply +28 VDC to the appropriate terminals (see Appendix A) to change from position 1 to position 2. Point the antenna towards the sky at some angle and observe the integrated video voltage output. If it is drifting too much, leave some more time for warm-up and see whether there is any improvement in the output stabiity. If it is not within the expected acceptable limits, reduce the IF amplifier gain slightly by turning the gain control potentiometer in a counter-clockwise direction. A full revolution of the potentiometer changes the gain by as much as 20 dB, therefore, a fraction of the full revolution should be attempted at a time. Note that the output voltage will change and so does the offset

requirement. Now place the eccosorb in front of the antenna and note the output voltage. From this data, calculate the output slope, then trim the video amplifier gain (increase) slightly to get the desired slope. Repeat this until the expected stability is reached. At this point, adjust the offset using R3. The receiver is now ready for measurements.

For satisfactory operation, calibration must be performed as often as possible at regular intervals.

6.0 COMMENTS ON THE SYSTEM

6.1 Antenna

The horn antenna used in this system exhibits very high sidelobe levels, leading to an antenna temperature T_a consisting of the contribution from the mainlobe as well as from the sidelobes. In simple terms, this means the received signal is not only from the observation area (covered by the main beam), but also from the large surrounding area (covered by many sidelobes to different extents), which is clearly an undesirable effect. In the case of an imaging system, this situation becomes even more undesirable because the sidelobes have the effect of overlapping the scene under observation. Hence, it is reasonable to use low sidelobe antennas like a reflector antenna, an array antenna or (better) a corrugated horn.

6.2 Low Noise Amplifier (LNA)

Recall that the low noise amplifier sets the lower limit of the overall system dynamic range because this amplifier is designed to have a lower noise figure (i.e., lower thermal noise floor) than the rest of the system that follows. Also, it is well known that the NF contribution of the subsystem

(excluding LNA) to the complete system (including LNA) will be less by a factor of the amplifier gain (actually, a little better than this - for exact expressions refer to Ulaby, Moore and Fung [1981]). It is for this reason the low noise amplifier, usually, is the first component in any receiver. An exception to this is that in high power radars (transmitting power, not the rec [illegible] echo signal) using a single antenna for transmission and reception, a pro[illegible] device called a "limiter" (sometimes two of them: one a tube type and the other a solid-state type) will be used before the amplifier to protect the sensitive receiver from large leakage power coupled from the transmitter. In radiometers, there is no need for any such protection, but a switch is usually required before the amplifier for calibration purposes. Also, the use of such LNA is always advantageous in radar if properly selected, whereas in radiometers it has demerits as well.

<u>Merits</u>:

1. A smaller noise figure, lower thermal noise floor, thus a better sensitivity.
2. Allows the use of slightly inferior components in the rest of the front end.

<u>Demerits</u>:

1. If not properly selected, the upper limit of the system dynamic range may be degraded.
2. The foremost disadvantage is the large gain variations associated with these.
3. They are more expensive than any other component in the system.
4. Above millimeter frequency range, no solid-state LNAs are currently available in the market.

In light of the above discussion, for this particular system it may be reasonable to eliminate the LNA, if a good quality mixer (low conversion loss and high LO to RF isolation, among other things) is available for use.

6.3 Mixer-Preamplifier

The mixer currently used has very bad high single sideband (SSB) conversion loss and very low LO to RF port isolation. The undesirable effect of the low isolation is not only that some of the signal is lost in multiple reflections, but also that the variations in the local oscillator noise contribution will appear as a signal at the receiver output leading to erroneous data. Hence, the system performance can be improved by using a better quality mixer.

Currently no preamplifier is used in this system. But a mixer integrated with a preamplifier capable of a combined gain (preamp gain-mixer conversion loss) of about 30 to 40 dB is desirable.

6.4 IF Amplifier

Assuming that all of the other parameters remain the same, the radiometric resolution could be improved (reduced) by increasing the bandwidth (see Section 2.0). For a fixed-time-bandwidth product the integration time can be reduced by increasing the bandwidth thus achieving not only the same resolution, but also higher output data rates. Because of this, radiometers (unlike radars) always operate with a large 3 dB bandwidth. In this case a 3 dB bandwidth of 8 MHz is not only reasonable but also quite practical, as many manufacturers supply 30 MHz IF amplifiers with as much as 10 MHz bandwidths.

6.5 Square-Law Detector

Proper selection of this device is critical for radiometers. Some of the important features to be considered are: dynamic range, linearity of the transfer characteristic, sensitivity, and stability. The first three can be satisfied by most of the detectors on the market, whereas the last one limits the choice of detectors. Crystal detectors have good sensitivity but, in terms of stability, they are not the best. Tunnel diode detectors are known to demonstrate almost an order better stability compared to crystal detectors, but with reduced sensitivity. While the reduction in sensitivity can be compensated for elsewhere in the design, the stability cannot; therefore, it is much better to use a tunnel diode detector for square-law detection.

6.6 Video Amplifier

The fact that the temperature coefficient for various passive components (resistors and capacitors) used in this system can cause remarkable changes in the resolution of the system is important. Hence, it is reasonable to use very low tolerance (may be 1%) components.

7.0 SUMMARY

A detailed and systematic design procedure for a practical 13.9 GHz radiometer has been presented. Various design tradeoffs were discussed and the effect of various parameters on the feasible resolution of the total power radiometer was shown. Required modifications were spelled out for converting this unit into a switching-type radiometer. To make the task of trouble-shooting easier, some of the important measurements were also discussed. A simple calibration procedure was used to illustrate the principles. Many of

the difficulties with the components used in the current system were made clear and, in turn, some suggestions were prescribed to improve the overall receiver performance.

REFERENCES

Evans, G. and C.W. McLeish, RF Radiometer Handbook, Artech House, 1977.

Laverghetta, T.S., Handbook of Microwave Testing, Artech House, 1981.

Ulaby, F.T., R.K. Moore and A.K. Fung, Microwave Remote Sensing Active and Passive, Volume 1, 1981; Volume 3, 1984, Addison-Wesley Publishing Company.

APPENDIX A

Standard Gain Horn

Model number	12-12
Nominal gain	24.7 dB
Nominal bandwidth	9° E-plane
	10° H-plane
Directivity	26.23 dB
Radiation efficiency	70.3% at 13.9 GHz

Figure A-1 shows the dimensions of the antenna.

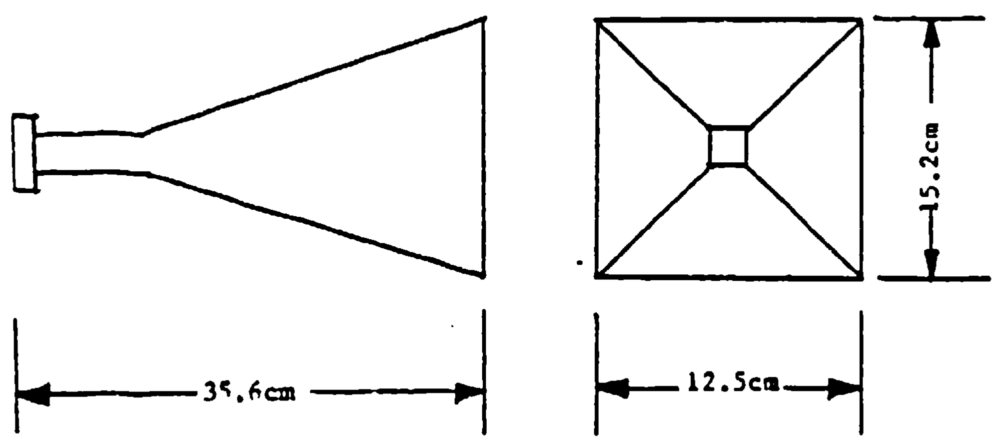
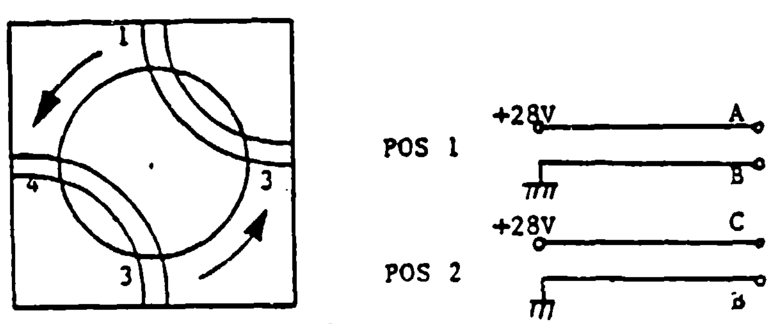


FIGURE A.1: Antenna Dimensions

Waveguide Switch

Model number	82152-33D00200
Insertion loss	0.1 dB
Isolation	60 dB
VSWR	1.1:1
Actuator voltage	+28 V
Current	1A max.
Wiring:	
Inside the switch	A = Black
	B = Orange
	C = White
Outside the switch	A = Black
	B = Red
	C = White

Figure A-2 shows the switch positions and the connections required to change from one position to the other.



Latched in position 1

FIGURE A.2: Waveguide Switch Details

Tunnel Diode Amplifier

Model number	T8503
3 dB bandwidth	13.4 - 14.4 GHz
Nominal gain	16 dB
Noise figure	4.85 dB
VSWR	1.1:1
Power output at 1 dB GCP	-4.0 dBm
Supply voltage	-15 VDC
Current	15 mA

No adjustments are required.

Bandpass Waveguide Filter

Bandwidth	50 MHz
Insertion loss	0.3 dB
Stop band attenuation	> 50 dB (100 MHz away from 3 dB point)

Tuned for maximum receiver gain

Crystal Oscillator

Model number	SSX-0225 (EI)
Frequency	13.929 GHz
Power output	6.2 dBm (measured at the output of the isolator)
Supply voltage	-20.0 VDC
Current	< 300 mA

Mixer

Model number	1210-2
DSB noise figure	17.2 dB
LO-RF isolation	8.2 dB
VSWR at the RF port	1.6:1 at 13.9 GHz

IF Amplifier

Model number	EVT30LL37
Center frequency	30 MHz
3 dB bandwidth	1.2 MHz
Noise figure	2.0 dB
Maximum power gain	82 dB
Saturated power output	+22 dBm
Supply voltage	+15 VDC at 170 mA
	-15 VDC at 30 mA

Crystal Detector

Model number	HP 423B
Sensitivity	500 mV/mW
Dynamic range	60 dB

Video Amplifier

Nominal gain	60 dB
Integrator time constant	.893 sec
Input impedance	1K ohm

APPENDIX B

```
      TAU = KA * DELTA
      TAUOZ = TAUOZ + 0.5*(TAU + TAUP)
      TAUP = TAU
      IF(Z.LT.H+.0001.OR.Z.GT.H-.0001)TAUOH=TAUOZ
      CONTINUE
   20 SUM = SUM + KA*T*EXP(-1.*TAUOZ*SEC)*C
C          COMPUTE Tdown AND Tsky
      TDN = SUM * DELTA * SEC
      TSKY = TDN + 2.7 * EXP(-1.*TAUOZ*SEC)
      TNAT=TAUOH*T0+(1.-TAUOH)*(1.-CSP)*TC+(1.-TAUOH)*CSP*TSKY
   97 RETURN
  300 FORMAT(1X,'ALT    KO2        KH2O     K      WEIGHT')
  301 FORMAT(1X,F4.1,4F8.5)
      END
C          SUBROUTINE H2OVAPOR
      SUBROUTINE H2OVAPOR(F,P,T,R,KH2O)
      REAL KH2O
      G = 2.35*P/1013.*(300./T)**0.626*(1.+0.018*R*T/P)
      KH2O = 2.*F*F*R*(300./T)**1.5*G*(300./T*EXP(-644./T)/((494.4
     &-F*F)**2+4.*F*F*G*G) + 1.2E-06)
      RETURN
      END
C          SUBROUTINE OXYGEN
      SUBROUTINE OXYGEN (F,P,T,R,KO2)
      REAL KO2, FN(20,2), YN(20,2)
      COMMON /BLOCK1/FN, YN
C          INITIALIZE THE CONSTANTS
      GN = 1.18 * P/1013. * (300./T)**0.85
      GB = 0.49 * P/1013. * (300./T)**0.89
      SUM = 0.
C          PERFORM SUMMATION OVER N
      I = 0
      DO 10 N=1,39,2
      I = I + 1
      PHIN = 4.6E-03 * 300./T * (2.*N + 1.) *
     &EXP(-6.89E-03 * N * (N+1.)*300./T)
      DNP = SQRT(N * (2.*N + 3.)/((N + 1.)*(2.*N + 1.)))
      DNN = SQRT((N + 1.)*(2.*N - 1.)/(N*(2.*N + 1.)))
      GN1 = (GN*DNP*DNP + P*(F-FN(I,1))*YN(I,1))/
     &((F-FN(I,1))**2 + GN*GN)
      GN2 = (GN*DNP*DNP + P*(-1.*F-FN(I,1))*YN(I,1))/
     &((-1.*F-FN(I,1))**2 + GN*GN)
      GN3 = (GN*DNN*DNN + P*(F-FN(I,2))*YN(I,2))/
     &((F-FN(I,2))**2 + GN*GN)
      GN4 = (GN*DNN*DNN + P*(-1.*F-FN(I,2))*YN(I,2))/
     &((-1.*F-FN(I,2))**2 + GN*GN)
   10 SUM = SUM + PHIN * (GN1 + GN2 + GN3 + GN4)
C          COMPUTE F' & KO2
      FPRIME = 0.7 * GB/(F*F + GB*GB) + SUM
      KO2 = 1.61E-02 * F*F * FPRIME *P/1013. * (300./T)**2
      RETURN
      END
C          SUBROUTINE ATMOS
      SUBROUTINE ATMOS(Z,P,T,R,P0,T0,R0)
C          COMPUTE THE TEMPERATURE AT ALTITUDE Z
      T11 = T0 - 6.5*11.
      T = T11
      IF (Z.LT.11.) T = T0 - 6.5*Z
      IF (Z.GT.20.) T = T11 + Z - 20.
C          COMPUTE THE PRESSURE AT ALTITUDE Z
      P = P0 / EXP(Z/7.7)
C          COMPUTE THE WATER-VAPOR DENSITY AT ALTITUDE Z
```

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```
C         MAIN CALLING ROUTINE
      REAL FN(20,2), YN(20,2)
      INTEGER POL
      COMPLEX EW
      COMMON /BLOCK1/FN, YN
      REAL KH2O, KO2
      DATA (FN(I,1),I=1,20)/56.2640,58.4466,59.5910,60.4340,
     &61.1506,61.8002,62.4112,62.9980,63.5685,64.1273,64.6797,
     &65.2241,65.7647,66.3020,66.8367,67.3694,67.9007,68.4308,
     &68.9601,69.4887/
      DATA (FN(I,2),I=1,20)/118.7503,62.4863,60.3061,59.1642,
     &58.3237,57.6125,56.9682,56.3634,55.7838,55.2214,54.6711,
     &54.1300,53.5957,53.0660,52.5422,52.0212,51.5030,50.7073,
     &50.4735,47.7613/
      DATA (YN(I,1),I=1,20)/4.51,4.94,3.52,1.86,.330,-1.03,
     &-2.23,-3.32,-4.32,-5.26,-6.13,-6.99,-7.74,-8.61,-9.11,
     &-10.3,-9.87,-13.2,-7.07,-33.0/
      DATA (YN(I,2),I=1,20)/-0.214,-3.78,-3.92,-2.68,-1.13,
     &.344,1.65,2.34,3.91,4.93,5.04,6.76,7.55,3.47,7.01,10.3,
     &9.36,13.3,7.01,26.4/
C         SCALE THE YN'S PROPERLY
      DO 5 I=1,20
      YN(I,1) = YN(I,1) * 1E-04
    5 YN(I,2) = YN(I,2) * 1E-04
C         INPUT SEQUENCE
      IFILE = 02
      WRITE(06,201)
      READ(05,200) IX
      IF (IX-3) ,99,99
      IF (IX-1) 6,6,
      WRITE(06,202)
      GO TO 99
    6 WRITE(06,203)
      READ(05,200) P0,T0,R0
      IF (P0) ,,7
      WRITE(06,204) "PRESSURE"
      GO TO 6
    7 IF (T0) ,,8
      WRITE(06,204) "TEMPERATURE"
      GO TO 6
    8 IF (R0) ,,9
      WRITE(06,204) "WATER VAPOR"
      GO TO 6
    9 WRITE(06,205)
      READ(05,200) F
      IF (F) ,,11
      WRITE(06,204) "FREQUENCY"
      GO TO 9
   11 IF (F-100.) 10,10,
      WRITE(06,206)
      GO TO 9
   10 WRITE(06,210)
      READ(05,200) IX
      IF (IX-1) 12,12,
      PRINT,'ENTER POLARIZATION:1-H,2-V'
      READ 20,POL
   20  FORMAT(V)
      PRINT,'ENTER REAL & IMAG PARTS OF WATER DIEL CONST'
      READ 20,EW
      PRINT,'ENTER RADIOMETER ALTITUDE (Km) ABOVE WATER LEVEL'
      READ 20,H
   12 WRITE(06,207)
      READ(05,200) ANG
      IF (ANG) ,13,13
      WRITE(06,204) "ANGLE"
```

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```
      GO TO 12
 13   IF (ANG-80.) 14,14,
      WRITE(06,200)
      GO TO 12
 14   CALL TEMP(F,POL,EW,H,ANG,P0,T0,R0,TSKY,TWAT,TAU0Z,IFILE)
      WRITE(06,100) F,POL,EW,H,P0,T0,R0,TSKY,TWAT,TAU0Z
      WRITE(IFILE,100) F,POL,EW,H,P0,T0,R0,TSKY,TWAT,TAU0Z
      GO TO 12
 99   STOP
100   FORMAT(1X,'FREQ = ',F5.1,' GHz',
      /,1X,'Polariz (H=1,V=2) =',I3,'  Water diel const=',2F6.2,
      /,1X,'Altitude=',F3.3,' Km',
      /,1X,'Po =',F5.1,' mbar  To =',F5.1,' K  Ro =',F5.2,' g/m3',
      /,1X,'Tsky = ',F7.4,' K  Twater = ',F7.4,' K',
      /,1X,'OPTICAL DEPTH (Tau) =',F8.5,' Np')
200   FORMAT(V)
201   FORMAT(1X,/,1X,'ENTER 1 FOR STANDARD ATMOSPHERE',/,1X,
      '      2 FOR USER DEFINED ATMOSPHERE',/,1X,
      '      3 TO EXIT PROGRAM')
202   FORMAT(1X,'MODIFY PROGRAM FOR USE WITH NEW ATMOSPHERE MODEL')
203   FORMAT(1X,'ENTER SURFACE PRESSURE (mbars), SURFACE TEMPERATURE',
      ' (Degrees Kelvin),'/,1X,' AND SURFACE WATER VAPOR DENSITY ',
      '(g/m3)')
204    FORMAT(1X,/,'INVALID',A11,'ENTRY',/,1X,'--- TRY AGAIN')
205   FORMAT(1X,'ENTER FREQUENCY (GHz)')
206   FORMAT(1X,'FREQUENCY SHOULD NOT EXCEED 100 GHz',/,1X,
      '--- PLEASE TRY AGAIN')
207   FORMAT(1X,'ENTER ANGLE FROM ZENITH (Degrees)')
208   FORMAT(1X,'ANGLE IS NOT TO EXCEED 80 Degrees',/,1X,
      '--- PLEASE TRY AGAIN')
210   FORMAT(1X,"ENTER 1 FOR Tsky COMPUTATION",/,1X,
      "      2 FOR Twater CALCULATION AT A GIVEN ALTITUDE")
      END
C           SUBROUTINE TEMP
      SUBROUTINE TEMP(F,POL,EW,H,ANG,P0,T0,R0,TSKY,TWAT,TAU0Z,IFILE)
      REAL KO2,KH2O,KA
      INTEGER POL
      COMPLEX EW
C           INITIALIZE THE CONSTANTS
      PI = 4.*ATAN(1.)
      SEC = 1./COS(ANG*PI/180.)
      N = 3200
      Z0= 32.
      DELTA = Z0/FLOAT(N)
      SUM = 0.
      TAUP = 0.
      TAU0Z = 0.
      WRITE(IFILE,300)
      CALL GAMA(POL,EW,H,ANG,GSP)
C           BEGIN INTEGRATION
      DO 20 I=1,N+1
      C = 1.
      IF (I.EQ.1.OR.I.EQ.N+1) C = 0.5
      Z = (I-1.)*DELTA
      CALL ATMOS(Z,P,T,R,P0,T0,R0)
      CALL H2OVAPOR(F,P,T,R,KH2O)
      CALL OXYGEN(F,P,T,R,KO2)
C           CONVERT FROM dB/km TO Neper/km
      KO2 = KO2 * 0.23026
      KH2O = KH2O * 0.23026
      KA = KO2 + KH2O
```

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```
      R = R0 / EXP(Z/2.)
      RETURN
      END
      SUBROUTINE GAMA(POL,EW,H,ANG,GSP)
      COMPLEX EW,A,B
      INTEGER POL
      AN= ANG*3.14159/180.
      A=CSQRT(EW-SIN(AN)*SIN(AN))
      B=CMPLX(COS(AN),0.)
      IF(POL.EQ.2) GOTO 10
C     HORIZONTAL POLARIZATION
      GSP=(CABS((B-A)/(B+A)))**2
      GOTO 20
C     VERTICAL POLARIZATION
 10    CONTINUE
       GSP=(CABS(((EW*B)-A)/((EW*B)+A)))**2
 20    RETURN
      END
```

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